

Compact, Frequency-Agile, Absorptive Bandstop Filters

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Abstract — Two new absorptive bandstop filter circuit topologies are introduced. The preferred topology is analyzed and its advantages over traditional approaches quantified. A corresponding varactor-tuned microstrip absorptive bandstop filter is described and its unique ability to maintain its attenuation while tuning across a broad frequency range using only resonant frequency tuning is demonstrated.

I. INTRODUCTION

Miniature electronically tunable bandstop filters are currently of interest for suppression of signal interference in receivers of frequency-agile communications systems [1]. In bandstop filters that reflect stopband signals, the unloaded Q 's (Q_u) of the filters' resonators limit both selectivity and ultimate stopband attenuation. And, since Q_u is generally proportional to resonator volume, the quest for greater selectivity and stopband attenuation is generally at odds with the drive towards miniaturization.

Bandstop filters that substantially absorb stopband signals, although still limited in performance by resonator Q_u , can be orders of magnitude more selective than reflective bandstop filters with identical resonator Q_u [2]. Absorptive bandstop filters function by matching the source resistance into the resonator resistances (resonator impedances at resonance) at stopband frequencies, so that signal power is dissipated in, rather than reflected from, the lossy resonators. Microwave absorptive bandstop filters have traditionally been comprised of lossy reflection-mode networks coupled to 3dB hybrid couplers [3] or circulators [4]. But, this dependence on hybrids or circulators is a significant practical constraint.

Recently, the author introduced the concept that absorptive reciprocal bandstop filters can be realized by any passive reciprocal linear network with multiple signal paths, multiple resonators, and at least one principally all-pass delay path [2]. In particular, it appears that any multiple-signal-path reflective bandstop filter topology of order $2n$ is capable of realizing an absorptive bandstop transmission characteristic of order n . While two specific topological examples were described in [2], they were not optimal from a standpoint of miniaturization or passband insertion loss. In this paper, two additional, more compact, absorptive bandstop filter topologies will be described.

Although sometimes a first-order bandstop response is all that is required, more often a higher-order response is desired. Absorptive bandstop filters can be intrinsically higher-order, as in [4], but such filters are not very amenable to being frequency agile due to the large number of frequency sensitive inter-resonator couplings. This paper anticipates realizing a higher-order response by cascading lower-order absorptive bandstop sub-networks.

Also, prior electronically-tunable passive reflective bandstop filters, such as in [5], have suffered appreciable performance degradation over the frequency tuning range due to frequency dependent loss in the tuning elements and resonators, as well as frequency dependent phase shift in the transmission lines used to couple the resonances. In contrast, the new absorptive bandstop filter topologies are ideally suited to electronic frequency control, with the unique ability to partially compensate for frequency dependent losses, couplings, and phase shifts by adjustments to resonant frequencies alone. A description of a first-order absorptive frequency-agile bandstop filter that was built and tested is provided as an example.

II. COMPACT ABSORPTIVE NOTCH FILTERS

Two new first-order, two-resonance, absorptive bandstop filter topologies are illustrated in Fig. 1. The preferred topology is that of Fig. 1(a). It is equivalent to the topology of the "second-order building block" in [6] and is essentially a second-order bandpass filter connected in parallel with an all-pass nominally-90°-phase-shift element. It may also be described as a generalized version of the "cross-coupled Brune section" in [7] and [8]. Fig. 1(b) is equivalent to the topology of the "Modified Doublet" in [9] and is essentially a parallel connection of two first-order bandpass filters and an all-pass nominally-90°-phase-shift element. Although the circuit of Fig. 1(b) has the same phase shift element and number of resonances as the circuit of Fig. 1(a), it lacks the other's symmetry, convenience of having all resonant elements on one side, and obvious compatibility with dual-mode resonator technologies (which is used to good advantage in [10]). The remainder of this paper focuses on the topology of Fig. 1(a).

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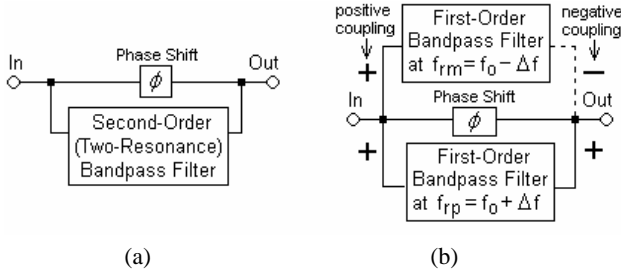


Fig. 1. Conceptual diagrams of first-order absorptive bandstop filters based on (a) a single second-order bandpass filter and (b) two first-order bandpass filters.

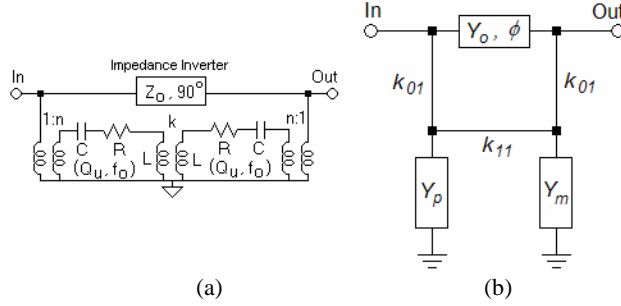


Fig. 2. Representative circuit schematics of Fig. 1(a) based on (a) an ideal-transformer-coupled lumped-element bandpass filter with mutual-inductance coupled resonators tuned to the same frequency and (b) a generalized admittance-inverter-coupled Brune section with potentially different admittances Y_p and Y_m .

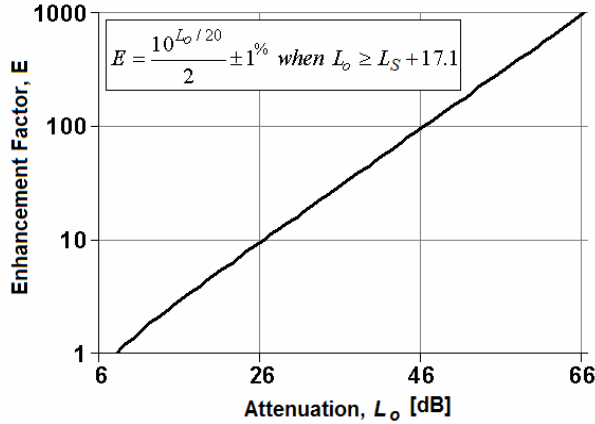


Fig. 3. Graph of fractional-bandwidth enhancement factor versus minimum attenuation at the notch center frequency.

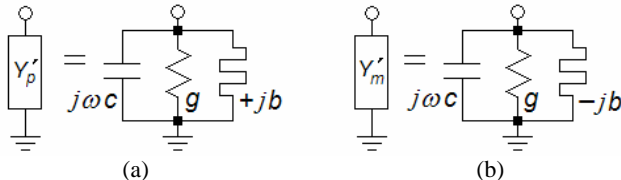


Fig. 4. First-order highpass-prototype admittance definitions for (a) Y_p and (b) Y_m in Fig. 2.

A. Generalized Cross-Coupled Brune Section Theory

When a second-order bandpass filter is connected in parallel with an all-pass phase shift element, as in Fig. 1(a), a particularly compact form of absorptive bandstop filter can be realized. A representative circuit schematic of such a filter is shown in Fig. 2(a), and is comprised of two lumped-element resonators that are coupled by transformers of turns ratio $1:n$ to an impedance inverter of impedance Z_o and nominal electrical length of 90° at resonant frequency f_o . The resonators are coupled to each other by a mutual inductance $m=kL$, where k is the coupling coefficient, and have identical Q_u , $Q_u=2\pi f_o L/R$. A minimum fractional 3dB stopband bandwidth of about

$$bw_{3dB} = 2/Q_u, \quad (1)$$

an infinite attenuation at f_o , and an essentially infinite return loss at all frequencies are realized by choosing

$$n = \sqrt{2R/Z_o}, \quad k = 1/Q_u, \quad \text{and} \quad Z_o = R_s = R_L, \quad (2)$$

where R_s and R_L are the source and load impedances. In this case, the general fractional bandwidth for a stopband band-edge attenuation value L_s (in dB) is approximately

$$bw = 2 / (Q_u \sqrt{10^{L_s/10} - 1}), \quad (3)$$

while, for a traditional first-order reflective bandstop filter with an attenuation L_o at center frequency f_o , it is [2]

$$bw_{trad} = \sqrt{10^{L_o/10} - 10^{L_s/10}} / (Q_u \sqrt{10^{L_s/10} - 1}). \quad (4)$$

Thus, a fractional-bandwidth (or selectivity or effective Q_u) enhancement factor, graphed in Fig. 3, is defined as

$$E = bw_{trad}/bw = \sqrt{10^{L_o/10} - 10^{L_s/10}} / 2. \quad (5)$$

The circuit of Fig. 1(a) can also be represented by the cross-coupled Brune section of [7] or [8], which is shown in a generalized (and not necessarily first-order) form in Fig. 2(b). Ideal admittance inverters k_{oi} couple generalized one-port networks (which could be like those in [4]), with potentially differing driving-point admittances, Y_p and Y_m , to the ends of a phase shift element of admittance Y_o and frequency-dependent phase shift ϕ , while admittance inverter k_{11} directly couples Y_p to Y_m .

To better understand the behavior of a first-order version of this absorptive bandstop filter it is most convenient to work with the high-pass prototype, with infinite attenuation at radian frequency $\omega=0$. The first-order highpass prototype is represented by Fig. 2(b), with $Y_p=Y_p'=g+j(\alpha c+b)$ and $Y_m=Y_m'=g+j(\alpha c-b)$ as shown in Fig. 4, where b is a frequency invariant susceptance, g is a conductance, and c is a capacitance. It will have $|S_{21}(j0)|^2=0$, be fully absorptive ($|S_{11}(j\omega)|^2=0$), and have exactly the minimum bandwidth given in (1) and (3), provided that

$$\phi = \pi/2, \quad Y_o = R_s = R_L, \quad b = 0, \quad k_{11}=g, \quad \text{and} \quad k_{oi}^2 = 2g. \quad (6)$$

Surprisingly, if $\phi = \pi/2$, $Y_o = R_S = R_L$, and $b=0$, the impedance match is entirely independent of the admittances Y_p' and Y_m' (which is a useful property for switched bandstop filter applications)! And, for any arbitrary attenuation at $\omega = 0$, the prototype has a perfect impedance match at all frequencies as long as

$$k_{01}^2 = 2 k_{11}, \quad (7)$$

again provided that $\phi = \pi/2$, $Y_o = R_S = R_L$, and $b = 0$.

On the other hand, for an arbitrary impedance match, the prototype has infinite attenuation at $\omega = 0$ as long as

$$k_{01} = \sqrt{Y_o \frac{b^2 + g^2 + k_{11}^2}{k_{11} \sin \phi}}. \quad (8)$$

Note that (8) guarantees infinite attenuation at $\omega=0$ for asymmetric, as well as symmetric, responses. Also, it is apparent from (8) that b can be used to adjust for changes in g , k_{11} , ϕ , k_{01} , and Y_o that might occur due to changes in operating environment or requirements. When the prototype is transformed to a bandstop filter, Y_p' and Y_m' transform into resonators and b effectively becomes a frequency offset between the resonators. Hence, it should be possible to tune the two resonators to frequencies offset above and below the notch center frequency to maintain notch attenuation while other parameters are changing.

III. EXPERIMENTS

A varactor-tuned microstrip absorptive bandstop filter of the type in Fig. 2 was constructed as shown in Fig. 5. A grounded Metelics MGV-125-20-E25 varactor is attached to one end of each of the two open-ended microstrip resonators to independently control their resonant frequencies. The varactors are quoted as having a capacitance range of 1.5 to 0.1 pF for 0.5 to 20 V reverse bias voltage (V_{rb}) and a 1.68Ω series resistance at $V_{rb} = 4V$. The $37 \times 70 \times 1.5$ mm substrate is Rogers RO4003 with 0.034 mm thick copper, 3.38 dielectric constant, and 0.0021 dielectric loss tangent. The microstrip resonators are electrically a quarter-wavelength long at 1.44 GHz.

A first set of measured responses for a bias voltage tuning range of 1V to 18V is shown in Fig. 6. A 20 dB bandstop attenuation depth was maintained over approximately a 30% tuning range and fractional bandwidth enhancements of a factor of 4 to over 100 were observed relative to *non-tuned* traditional notch filter approaches over the entire tuning range. All this despite the fact that measurements showed one resonator with a 12% to 344% better Q_u than the other in an amount that varied with varactor bias across the tuning range.

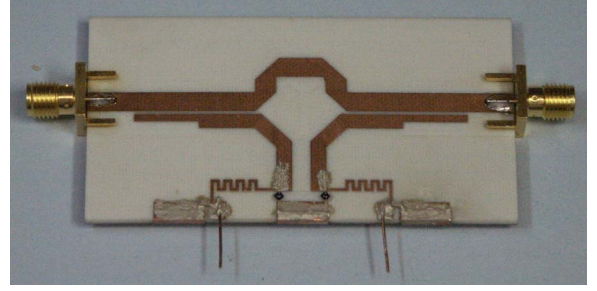


Fig. 5. Varactor-tuned microstrip absorptive bandstop filter.

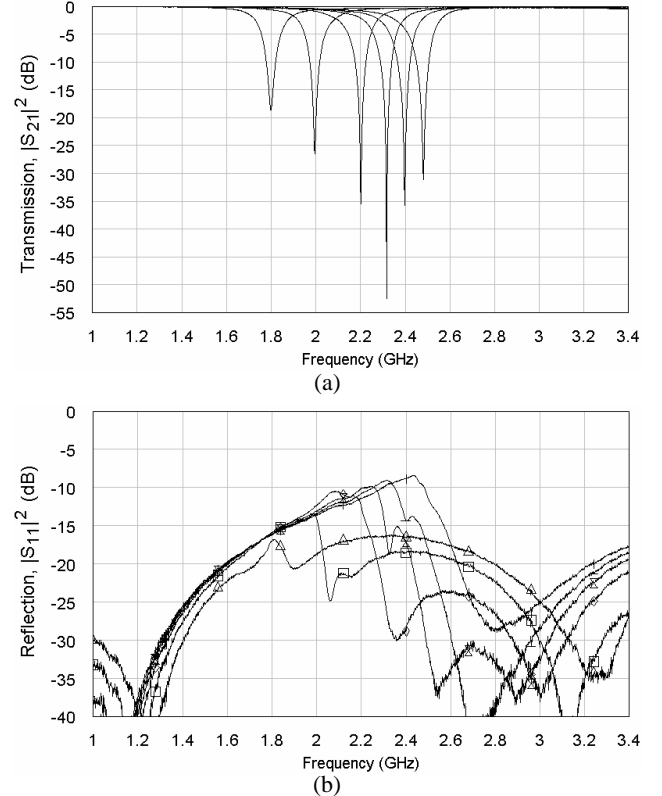


Fig. 6. Measured (a) transmission and (b) reflection responses of the experimental filter with a minimum 1V bias corresponding to the left-most curves and a maximum 18V bias corresponding to the rightmost curves.

It was found necessary to tune the parallel-line resonator-to-transmission-line couplings with metalized dielectric overlays prior to sweeping the varactor bias voltages. The relative varactor biases were then experimentally optimized to achieve the maximum bandstop attenuation at each center frequency in Fig. 6(a).

Another set of measurements, shown in Fig. 7, were obtained by first experimentally modifying the overlays to adjust the resonator-to-transmission-line couplings and then experimentally optimizing the relative varactor biases to maintain a minimum 15 dB return loss and a minimum 20 dB attenuation across a 10.5% tuning range.

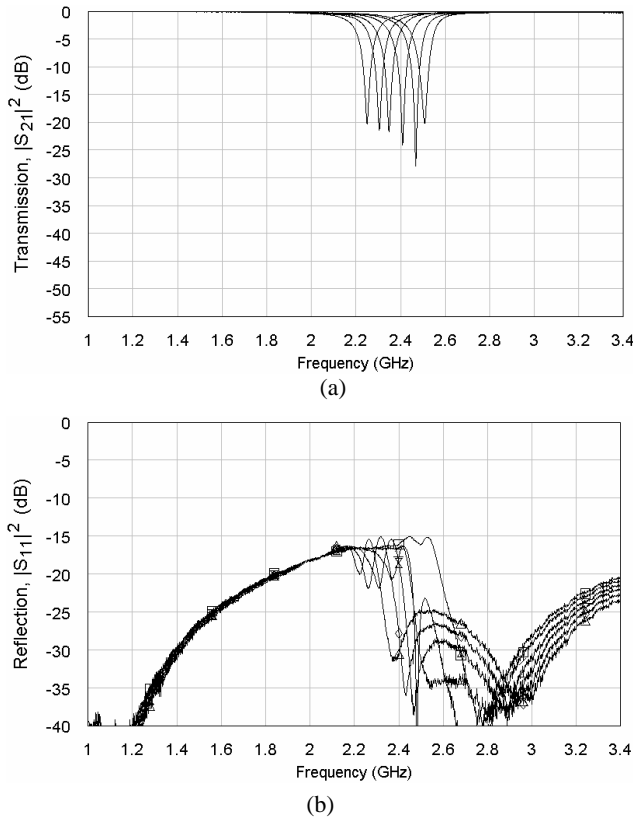


Fig. 7. Measured (a) transmission and (b) reflection response of the experimental filter, optimized for a 15dB impedance match across the entire tuning range, with a minimum 4V bias corresponding to the left-most curves and a maximum 20V bias corresponding to the rightmost curves.

IV. DISCUSSION

Simulated results exhibited far better performance than the measured results, in terms of tuning bandwidth and notch relative bandwidth and attenuation. The reasons for this are thought to be larger varactor series resistance than anticipated (especially at low bias voltages), excessive inter-resonator coupling due to an inadequate varactor grounding strategy, and inadequate consideration of the ground via inductance of both the varactors and bias capacitors. Also, while the simple first-order varactor bias circuit (a shunt 20 pF capacitor followed by a meandered quarter-wavelength 120 Ω microstrip line connected to the varactor cathode at the end of each microstrip resonator) proved sufficient in the circuit simulations, in retrospect a higher-order bias circuit is more prudent when attempting to realize notch attenuations in excess of 40 dB.

Offset tuning of resonator frequencies (i.e., increasing b) can only increase the right-hand side of (8). When maintaining attenuation is of primary concern, filters

should be designed such that (8) is satisfied when $b=0$ and the other parameter values represent the “worst case” situation over the desired tuning range, so that increasing b can compensate for anticipated changes in those values.

V. CONCLUSION

Specific new absorptive bandstop filter topologies have been introduced and a theory has been presented for the preferred topology. The theory suggests that the new topology might offer unique advantages for frequency agile bandstop filter applications. Experiments were done to gain additional understanding of the unusual filter in the context of an electronic frequency control requirement. The results offer encouragement that low-cost miniature bandstop filters capable of maintaining their performance over a broad tuning range are possible.

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